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MEMORANDUM

From: Al Gordon, Code 5211
To: Paul Heckman, Code 5211

Subj: Considerations concerning an active electromagnetic pipeline following system

Ref: (a) University of New Hampshire, Sonatrack I, The Design and Development of a Pipeline Survey Vehicle, by B. Haley 1976-1977.
(b) Innovatum, Inc. ltr dated 8 November 1975 from Paul A. Cloutier to Mr. Donald M. Rosencrantz.
(c) Geometrics, Applications Manual for Portable Magnetometers, by S. Breiner, 1973.
(d) Proceedings, 1978 Offshore Technology Conference, "Inspection of Buried Pipelines by Submersibles - Pipe Tracking and Pipe Logging Instrumentation," by Y. Durand and Alain Stankoff, pp. 207-216.
(e) NUC memo 6512/137-75 dated 22 August 1975 "Feasibility Investigation of using some form of metal detector as a sensor in a free-swimming, underwater object inspection vehicle," by M. Wolff
(f) Handbook of Electromagnetic Propagation in Conducting Media, by Martin B. Kraichman, U.S. Government Printing Office, 1970.
(g) Galejs, J., "Admittance of Insulated Loop Antennas in a Dissapative Medium," IEEE Trans. Antennas Propagation, Vol AP-3, March 1965, pp. 229-235.
(h) Ramo, Whinney, and Van Duzer, Fields and Waves in Communcations Electronics, Wiley, New York, 1967, p. 295
(i) Eshbach, Handbook of Engineering Fundamentals, 3rd Edition, Wiley, 1975, p. 1443
(j) Arnold Engineering Co., Tape Wound Cores, Publication TC-101B, Marengo, Illinois, 1972.
(k) American Institute of Physics, American Institute of Physics Handbook, Third Edition, McGraw-Hill, 1972, p. 5-247.
(l) Naval Undersea Center TP 294, Undersea Detection of Various Signals, "Static Electric and Magnetic Fields," by A. Gordon, July 1972, p. 17

1.0 INTRODUCTION

One of the missions which appears attractive for the NOSC free swimmer is pipeline inspection. With an ever-greater number of pipes distributing oil and gas from offshore wells, there is a need to insure that these pipes are not leaking and show no signs of immenent failure. A typical scenario would be the launch of the free swimmer some distance from the pipeline with the free swimmer programmed to cross the pipeline at some preset altitude. Aboard the free swimmer will be some sort of pipeline detection system, the subject of this memorandum, which will alert the onboard microprocessor that the pipeline has been crossed. The vehicle's control logic will then take over piloting and greatly influenced by the perm of the pipeline (reference (c)). This would appear to make the control problem more difficult for our free swimmer in a pipeline following/inspection mission. This same point has been made by Mannfred Wolf in reference (e). For this reason, this memorandum will not investigate passive DC systems further. It should be remembered, though, that these systems will have a detection advantage over active systems, and also that future, more detailed investigations may reveal the control problem to be soluble.

By using an active electromagnetic system, the control probelms are in principle more tractable since a simple dipole source is induced on the pipeline. Such a system has been build by Intersub-Development (reference (d)) and Shell (reference (e)). The Shell "Shellfish" system was designed as a towed fish to detect pipelines as it was towed over them. It operated at 50 Hz, and claimed the ability to detect an 8-inch pipeline buried 9 feet deep. Apparently this system was made of a single source and receiver and therefore would not be able to discriminate the pipeline's direction from a vehicle. The Intersub system overcomes this drawback by using one transmitter coil and two laterally separated receiver coils. From the drawing in the article, these coils are solenoidal (probably filled) and arranged so that the transmitter coil's axis is perpendicular to the two parallel receiver coils. This system is claimed to have a standoff distance of 3 meters against a 32-inch pipeline, whether buried or exposed, and to provide sufficient control for manual or automatic piloting.

Because the active electromagnetic systems provide good discrimination against buried objects and provide simple and easily usable control information, the balance of this article will examine them exclusively.

3.0 MAGNETIC SEARCH EQUATIONS

3.1 Nature of the Effect

In searching for a pipeline, we are dealing with an object which is both permeable and conductive. These properties give rise to two separate responses when the object is illuminated by an AC magnetic field. First, since the object is permeable, the magnetic field tends to induce a dipole moment parallel to the incoming field by magnetizing the object. At the same time, since the object is conductive, eddy currents are set up. These currents, by Lenz' law set up a dipole moment anti-parallel to the incident field. The two effects, magnetization and eddy currents, have opposite frequency behavior. Since eddy currents are created by induction, there are no induced eddy currents at DC. Eddy currents increase with increasing frequency until, at a frequency when the skin depth is considerable less than the conductor size, the maximum eddy current dipole moment, which is proportional to the illuminated objects volume, is reached.

The induced dipole moment due to magnetization behaves exactly the opposite. At DC it has its maximum value, which is also proportional to the volume of the object. This magnetic moment is maintained until a frequency is reached where the skin depth is small enough to prevent an appreciable fraction of the object from being magnetized (the condition is $\frac{\mu\delta}{R} \sim 1$ where μ is the relative permeability, δ the skin depth, and R a characteristic dimension of the object). As the frequency is raised still further, the magnetization effect drops to zero.

Because of the rather thick well thickness of the pipe, it is unlikely that eddy currents will be suppressed at any frequency above a few Hertz. Also, depending on the magnitude of the relative permeability, some magnetization will be induced into the thousands of Hertz. To go higher in frequency would result in an unacceptably low water skin depth. Thus, the observed induced field will probably be a combination of dipoles produced by eddy currents and magnetization. Since these dipole moments are oppositely directed, there will be some cancellation over either effect acting alone. For analysis purposes, we shall ignore this cancellation and treat the eddy current effect alone.

One interesting results of the above considerations is that it may be possible to choose the frequency to minimize interference effects due to the free swimmer. An AC coil mounted on the vehicle will induce eddy currents on the vehicle's frame. This can, of course, be minimized by choosing the location and orentation of the coil with care. However, the eddy currents may be additionally suppressed by choosing the frequency low enough. Since the pipeline is much larger than the tubing on free swimmer, it seems eddy currents could be preferrentially suppressed on the free swimmer relative to the pipeline. As we shall see, choosing a low frequency minimizes losses in the water and minimizes changes in an immersed coil's impedence relative to its in-air value. However, at too low a frequency, atmospheric and other interfering noise sources will propagate into shallow water.

3.2 Governing Equations

In this section, we shall derive the magnetic field at a point dipole transmitter due to an induced dipole located a distance away. We shall use the infinite medium approximation, which assumes that the free swimmer and pipeline are far from the surface or boundries. Of course, with the pipeline at the bottom, this assumption is violated. However, the effect of the bottom, with its lower conductivity, appears mainly to increase the magnitude of the resulting fields. Hence, our infinite medium results should be conservative.

We will first work out the results for the transmitting dipole's moment along the line joining the transmitting dipole to the pipeline. Along the axis of the transmitting dipole, the electric and tangential magnetic fields are zero and the radial magnetic field, H, given by (reference (f))

$$H = \frac{m \cos \theta}{2\pi r^3}\left(1+\beta r + j\beta r\right) e^{-\beta r} e^{-j\beta r} \tag{1}$$

where m is the transmitting dipole's dipole moment, r is the distance to the observation point (i.e. in our case the pipeline), is the angle between the dipole's axis, and the line drawn to the observation point and β is the reciprocal of the attenuation length δ, i.e.,

$$\beta = \frac{1}{\delta} = \left(\frac{\omega \mu_0 \sigma}{2}\right)^{1/2} \tag{2}$$

where ω is the angular frequency (i.e. $\omega=2\pi f$), μ_0 is the permeability of free space, and σ is the conductivity of ocean water. Note that a time factor of $\exp[i\omega t]$ has been suppressed.

As mentioned above, the effect of the time verging magnetic field is to induce a magnetic moment in the pipe whose magnitude is proportional to incident magnetic field (see Appendix B, reference (f)), i.e.

$$m_i = \alpha_T H \tag{3}$$

where m_i is the magnetic moment induced on the pipeline (due to eddy currents) and α_T is the constant of proportionality. The subscript T indicates that the transmitting dipole moment is perpendicular to the pipe axis.

At the transmitting dipole, the magnitude of the induced field, H_i, is given by equation (1) with $m=m_i$ that is

$$H_i = \frac{m\alpha_T}{(2\pi)^2 r^6}\left\{(1+\beta r+j\beta r)^2 e^{-2\beta r} e^{-j2\beta r}\right\} \tag{4}$$

We have dropped the $\cos\theta$ dependence since we evaluated the field directly above the pipe. Labelling the term in brackets $G(\beta r)$, we note that G contains the entire frequency dependence. Expanding G into its real and imaginary parts, with $x=\beta r$, we have

$$\begin{aligned} \mathrm{Re}\, G(x) &= \left[(1+2x)\cos 2x + 2x(1+x)\sin 2x\right]e^{-2x} \\ \mathrm{Im}\, G(x) &= \left[2x(1+x)\cos 2x - (1+2x)\sin 2x\right]e^{-2x} \end{aligned} \tag{5}$$

Equations (3) through (5) are for a transmit coil whose dipole moment is oriented parallel to the line joining the dipole to the pipe. Another interesting geometry is for the transmitting dipole oriented perpendicular to both the pipeline's axis and the line joining the pipeline to the dipole. In the case of the transmitting dipole being a planar coil, this would correspond to the plane of the coil perpendicular to the bottom. For this geometry, the induced field at the dipole, H_i', is (from reference (f))

$$H_i' = \frac{m\alpha_T}{(2\pi)^2 r^6}\left\{\frac{1}{4}\left(1+\beta r+j(\beta r+2\beta^2 r^2)\right)^2 e^{-2\beta r} e^{-j2\beta r}\right\} \tag{6}$$

Note that the prefactor of the term in brackets is identical in both equations (4) and (6). Calling the term in brackets G' we have

$$\begin{aligned} \mathrm{Re}\, G'(x) &= \frac{1}{4}\left\{(1+2x-4x^3-4x^4)\cos 2x + (2x+6x^2+4x^3)\sin 2x\right\}e^{-2x} \\ \mathrm{Im}\, G'(x) &= \frac{1}{4}\left\{(2x+6x^2+4x^3)\cos 2x - (1+2x-4x^3-4x^4)\sin 2x\right\}e^{-2x} \end{aligned} \tag{7}$$

The above equations allow us to make some qualitative remarks concerning the behavior of the induced magnetic effect as a function of frequency and range. For a fixed number of attenuation lengths (i.e. βr=constant), equations (4) and (6) show that the magnitude of the effect drops as r^{-6}. This is true for localized objects, such as wellheads, where α_T is in depent of r. For extended objects, such as pipelines, it turns out that α_T is proportional to r (see next section) so the dependence on range is r^{-5}. In either case, this is an extremely rapid dropoff with range compared with, for example, the passive DC effect of a pipeline which decays as r^{-2}. It is this rapid dropoff with range that limits the AC induced effect to short ranges.

The frequency dependence is entirely contained within the functions G and G'. The easiest way to view the frequency dependence is to study the variation of G and G' as a function of $x=\beta r$, the number of attenuation lengths. Figure 1 shows $\|G(x)\|$, Re G(x) and Im G(x) as a function of x while Figure 2 shows the same information for G'. For G note that its maximum is reached at zero frequency where $\|G\|$ = Re G=1. Of course, our equations don't apply at zero frequency since the assumption of eddy currents as the dominant mechanism fails there. As the frequency is increased, the magnitude of the effect drops. However, ImG has an important maximum at $x=\beta r=1$, where it has a value of .594. ImG represents the quadrature returned signal; Figure 1 shows that by choosing βr properly (i.e. $\beta r \sim 1$) we can force the returned signal to be totally in quadrature with the transmitted signal without much loss in the magnitude of the effect. Such a choice could, in principle, allow discrimination

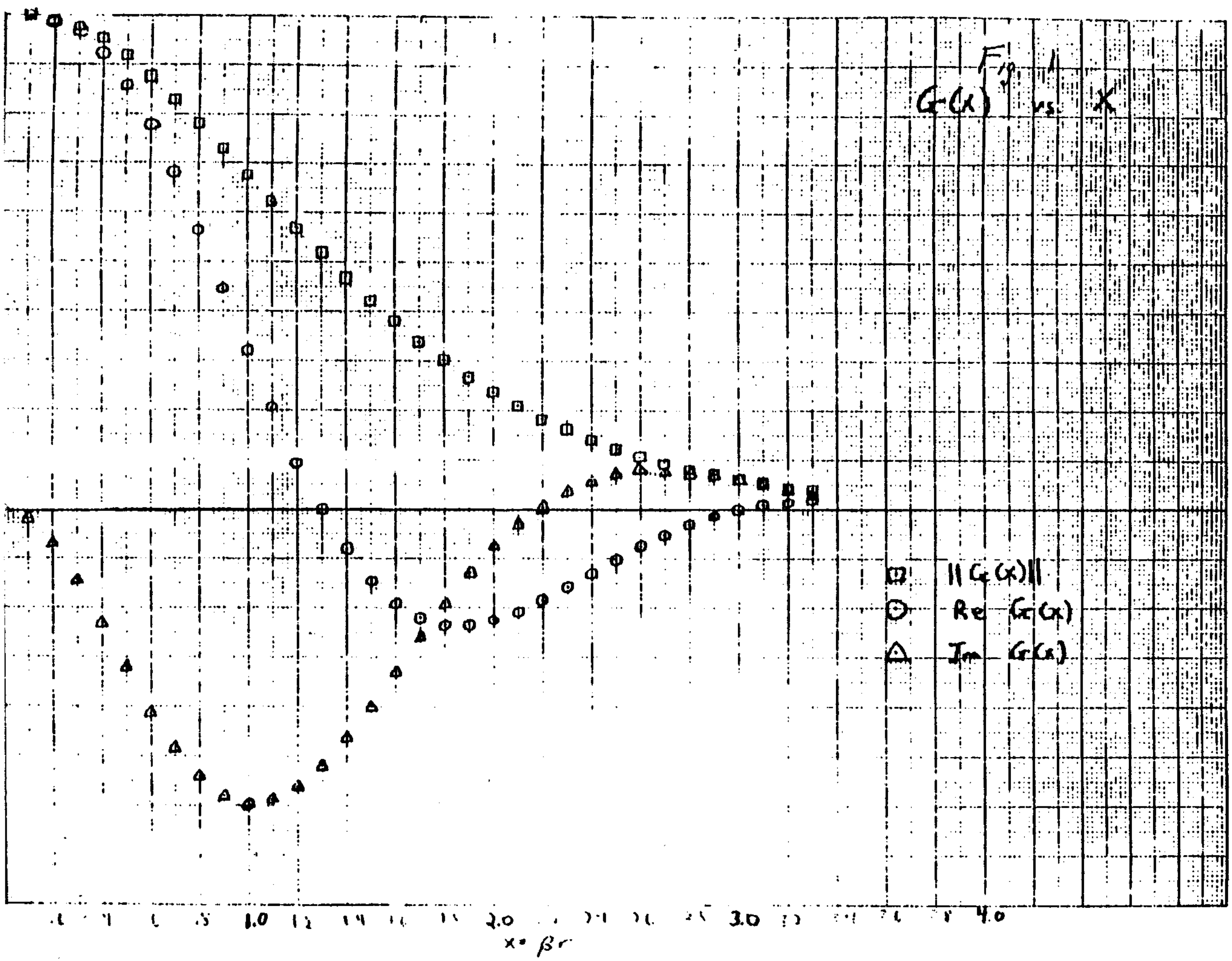
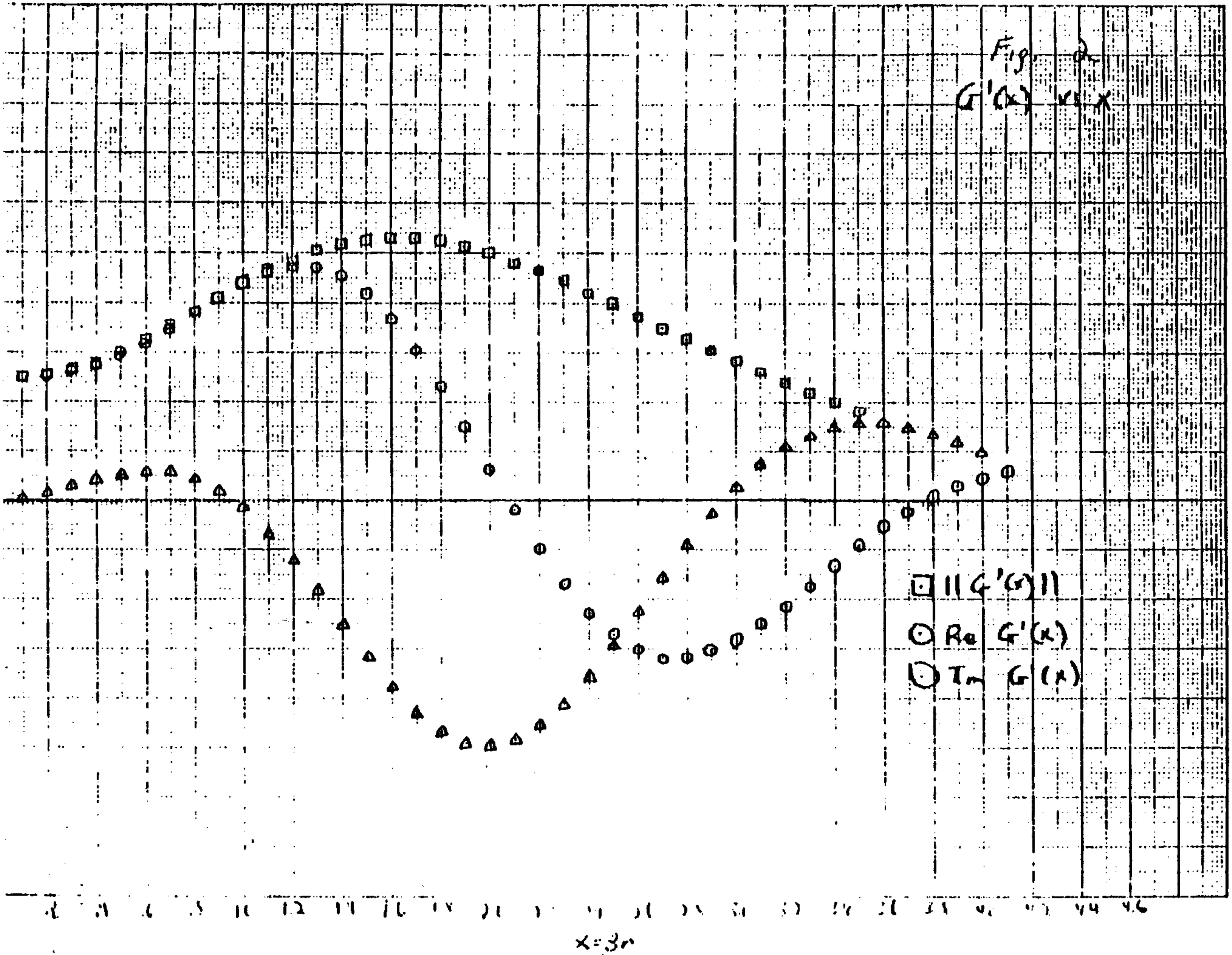


Fig. 1
G(x) vs. X
|| G(x) ||
Re G(x)
Im G(x)
1.0
2.0
3.0
4.0
x = βr

Fig. 2

$G'(x)$ vs x



between the desired return signal and the undesired transmitted signal. In any event, above $\beta r \approx 1$ G decays rapidly and is less than about .06 for $\beta r \geq 3$.

G' behaves rather differently than G. At low frequencies $\|G'\|$ is only .25. However, with increasing frequency $\|G\|$ increases reaching a magnitude of about .53 at $\beta r = 1.7$, after which it monotonically decreases. ImG' has a substantial maximum of about .5 at $\beta r = 2$. Although $\|G'\|$, just like $\|G\|$ becomes small at large βr, $\|G'\|$ decays much more slowly. For example, at $\beta r = 4$ $\|G'\|$ is still about .1. The reason for $\|G'\|$ decaying more slowly is that the electromagnetic field must become transverse at large βr so that $G/G' \to 0$ as $\beta r \to \infty$.

In summary, we have shown that AC-induced magnetic fields drop off quickly with r, the separation distance between the transmitting dipole and the illuminated object. For point objects this dependence is r^{-6}, for extended objects, such as pipelines the dependence is r^{-5}, as will be shown in the next section. The frequency dependence is more moderate, despite the appearance of the exponential factors $e^{-2\beta r}$ in equations (4) and (6). The effect can be made to be 60% of its greatest value at $\beta r = 1$, 50% at $\beta r = 2$ and is still 10% at $\beta r = 4$.

3.3 Induced Pipeline Magnetic Moment

In the last section, we used the fact from reference (f) that the induced magnetic moment of a metallic object is proportional to the incident magnetic field. This relationship is true when the size of the object is small compared with a wavelength in the surrounding medium, but large compared to a wavelength in the object. For cylinders, the constant of proportionality is given by (reference (f), appendix B)

$$\alpha_T = -nV \tag{8}$$

where V is the volume of the cylinder and n is a numerical factor satisfying $1 \leq n \leq 2$ and depending on the cylinder's length to diameter ratio.

In order to compute V, we must know the effective length of the pipeline. In the case of a long pipeline, this is considerably shorter than the physical length since substantial eddy currents are induced only on that portion of the pipeline immediately below the transmitting coil.

The most straight-forward way to estimate the effective pipeline length is to consider the induced moment to be composed of a linear dipole density, m(x), located along the pipeline axis. Let x=o be the point on the pipe axis directly below the transmitting dipole and a distance a from it. At point x on the pipe axis a dipole moment m(x)dx is induced. Disregarding for the moment the $\cos\theta$ factor the resulting induced field due to the induced moment m(x)dx varies as r^{-6}. At a distance x along the pipeline

$$r = (x^2 + a^2)^{1/2} \tag{9}$$

Thus, a small section of pipeline at x=o will produce an effect proportional to $(a)^{-6}$ while a small section at a distance x will have an effect proportional to $(a^2 + x^2)^{-6/2}$. If we define the effective half length, L, to be out to that point where the induced field at the transmitter is half that due to a point directly below then

$$\frac{a^6}{(a^2+L^2)^{6/2}} = \frac{1}{2} \tag{10}$$

Solving this equation we obtain

$$L = .51\, a \tag{11}$$

The effective length is 2L=1.02a, i.e. the length of the pipeline containing significant eddy currents is essentially equal to the distance from the pipe. In the following, we shall use L=1/2 a for simplicity.

In the limit of large length to diameter ratio, the value of n in equation (8) is 2 (reference (f), appendix B).

Using this value, we have

$$\alpha_1 = -2\,(2L)\,\pi\, a_p^2 = -2r\,\pi\, a_p^2 \tag{12}$$

where we have reintroduced the symbol r to represent the distance between dipole and pipeline and where a_p is the pipelines radius.

3.4 Dipole Moment to Produce a Given Field

By rearranging some of the previous equations, we can calculate the required dipole moment necessary to produce a given induced field, H_i at the transmitting dipole. Substituting equation (12) into equation (4) we have

$$H_i = \frac{-m\, a_p^2}{2\pi r^5}\, G\,(\beta r) \tag{13}$$

Rearranging and using $B = \mu_o H$

$$|m| = \frac{2\pi r^5 B_i}{\mu_o a_p^2\, |G(\beta r)|} \tag{14}$$

This gives the magnetic moment required for transmitting dipole whose moment is parallel to the line joining it with the pipeline. For the perpendicular case, equation (14) holds with G replaced by G'.

From equation (14) we can calculate the dipole moment required to receive a given magnetic field as a function of distance, pipe diameter, and frequency. As an example, let's evaluate this equation at a distance 4 meters from 54-inch pipeline (a_p=27 in=.686m), using a frequency such that βr=1, i.e. $\| G(1) \|$ =.58 (from figure 1). With $B = 1\gamma = 10^{-9}$ wb/m^2, we have

$$|m| = \frac{2\pi\,(4)^5\, 10^{-9}}{4\pi\, 10^{-7}\,(.686)^2\,(.68)} = 16\ \frac{\text{amp}\cdot\text{m}^2}{\gamma} \tag{15}$$

That is, at the transmitting dipole's location we need a magnetic moment of 16 amp - m^2 for every γ of induced field produced.

4.0 AC MAGNETIC SOURCES

In this section we shall review the applicable equations which describe coils and solenoids used as magnetic sources. These equations will relate the basic input parameters of the source such as size, weight, power to their resulting dipole moment, and associated electrical parameters (inductance, Q, etc).

4.1 Dipole Moment of a Circular Coil

Consider a planar circular coil of radius a carrying a current I. Let the coil consist of n turns of wire (wire radius = a_o) of resistivity ρ_1 and mass density ρ_m. Then this coil's dipole moment m, is given by

$$m = nI(\pi a^2) \tag{16}$$

The current is related to the real power input, P and the coil's resistance R by

$$I = \sqrt{\frac{P}{R}} \tag{17}$$

From (16) and (17) it is clear that to maximize the dipole moment for a given power input the coil's resistance must be minimized. The resistance R is the sum of the internal resistance R_i due to the wire's resistivity and R_e the external resistance due to conduction currents in the water. Assuming the wire used to make the coil is less than a skin depth in diameter then

$$R_i = \frac{n\rho_1 \ell}{a_o} \tag{18}$$

where $\ell = 2\pi a$ is the length of a single turn. The number of turns n may be eliminated from equation (18) by expressing it in terms of the coil's mass, M as

$$n = \frac{M}{\ell a_o \rho_m} \tag{19}$$

Substituting, we have

$$R_i = \frac{M\rho_1}{a_o^2 \rho_m} \tag{20}$$

The external resistance, Re, must be added to the internal resistance whenever the coil is surrounded by a conductor, such as seawater. Reference (f), Appendix C gives the value of this as

$$R_e = n^2 \frac{8}{3} \frac{\rho_2}{a} \left(\frac{a}{\delta}\right)^4 \tag{21}$$

where ρ_2 is the water's resistivity and δ the skin depth in water. Equation (21) only holds for $a \ll \delta$. For a on the order of δ/π, R_e will be higher unless the coil is insulated (reference (g)). Assuming this is done we substitute equations (17), (19) and (20) obtaining

$$m = \frac{a}{2}\left(\frac{P}{\frac{\rho_m \rho_1}{M} + \frac{2\rho_2 a}{3\pi^2 \delta^4}}\right)^{1/2} \tag{22}$$

It is interesting to note that the magnetic moment does not depend on the number of turns nor the size of wire. The choice of material is governed by the factor $\rho_m \rho_1$ which should be minimized. Of the common conductors aluminum, $\rho_m \rho_1 = 7.72 \times 10^{-5}\ \Omega\cdot Kg/m^2$ is superior to copper, $\rho_m \rho_1 = 15.41 \times 10^{-5}\ \Omega\cdot Kg/m^2$, producing about 40% more moment for the same size, weight and power input. Since for the diameters we'll be using, aluminum will represent a lot of wire (i.e. over three times as long as copper), we will base our numerical examples on copper.

It is interesting to compare the two terms in the denominator of equation (22). For copper $\rho_m \rho_1/M$ (M=10 lbs) equals $3.39 \times 10^{-5}\ \Omega/m^2$. For a coil radius of a=.25m and for δ=4m (i.e. f=3960Hz) the second term equals $1.65 \times 10^{-5}\ \Omega/m^2$. Thus the second term is about one half the first so that coils of this size weighing more than 20 pounds would not produce an appreciably larger moment at 4 KHz since the losses are dominated by seawater losses. For frequencies considerably lower than 4 KHz appreciably more than 10 pounds could be wound around a .25m radius coil and the magnetic moment would increase as the square root of the windings weight.

To complete our numerical example lets calculate the magnetic moment for a 10 pound, .25 m radius coil driven with 10 watts of power at 3960 Hz. According to equation (22),

$$m = \frac{(.25)}{2}\left(\frac{10}{(3.39\times10^{-5} + 1.65\times10^{-5})}\right)^{\frac{1}{2}} = 55.7 \text{ amp}\cdot m^2 \tag{23}$$

In equation (15), we found that it required a 16 amp-m^2 source dipole moment to yield an effect of one γ so with the coil configuration above we will see an effect of roughly 3.5 γ. For future reference, we note that at 40 Hz, the second term in the denominator of equation (23) vanishes and m=67.9 amp-m^2.

4.2 Electrical Design of a Coil

Continuing with the example of the last section we shall calculate the relavent electrical parameters pertainent to the employment of a coil. First we must choose s suitable wire size for the coil. If the wire diameter is too large there will be appreciable skin effect resistance which will lead to excess heating and inefficiency. Too small a gauge wire will result in increased stray capacitance and too many turns unnecessarily prolonging the winding process. For copper at 4 MHz, the skin depth is 1.05mm. We can make the wire 3-skin depths in diameter and only increase the DC resistance by about 15% (reference (h)). From standard copper wire tables (reference (i)), we see that #8 wire is almost exactly this diameter. Ten pounds of #8 wire corresponds to 200 feet and has a resistance of .126 ohm. At this resistance and ten watts we have a current of

$$I = \sqrt{\frac{P}{R}} = \sqrt{\frac{10}{.126}} = 8.92 \text{ amp} \tag{24}$$

This is considerably below the maxim current for #8 wire which is over 40 amps even for insulated wire. The number of turns is the total length divided by the circumference, i.e.

$$n = \frac{\ell}{2\pi a} = \frac{200 \cdot 12}{(2)(3.14)(.25)(39.34)} = 38.8 \text{ turns} \tag{25}$$

The inductance, L, of the coil is obtained from reference (h) as

$$L = n^2 a \mu_0 \left[\ln\left(\frac{8a}{a_B}\right) - 2 \right] \text{ henries} \tag{26}$$

where a_B is the radius of the bundle of wires. For a packing fraction P_f this equals

$$a_B = \frac{d_w}{2} \sqrt{\frac{N}{P_f}} \tag{27}$$

The largest packing fraction possible for circular wires is $\pi/4$. Using this value for our #8 wire, we have

$$a_B = \frac{.1285}{2} \sqrt{\frac{(38.8)(4)}{\pi}} = .451 \text{ in} \tag{28}$$

Using this value in equation (26), we find $L = 1.50 \times 10^{-3}$ Henrys.

At a frequency of 4 KHz, we have for the Q of the coil in air

$$Q = \frac{\omega L}{R} = \frac{(2\pi)(4\times10^{3})(1.5\times10^{-3})}{.126} \cong 300 \tag{29}$$

This is quite a large value of Q and probably will not be realized in practice due to stray capacitance, the finite insulation of the wires, less than ideal packing fraction, and presence of nearby objects such as the electrostatic shield.

To drive the above coil directly to the ten-watt level would require a source capable of driving an almost purely reactive load at 9 amps and an r.m.s. voltage V, given by

$$V = I X_L = I 2\pi f L = (9)(2)(\pi)(4000)(1.5\times10^{-3}) = 339 \text{ volts} \tag{30}$$

By resonating the load and making it purely resistive, we can considerably reduce the drive requirements. The capacitance, C, necessary to resonate an inductance, L, is given by

$$C = \frac{1}{(2\pi)^2 L f^2} \tag{31}$$

which for $L=1.5 \times 10^{-3}$ Henrys and f=4000 Hz yields C=1.05μf. Once resonated, the equivalent resistance, Req, of theparallel coil capacitor combination is given by

$$R_{eq} = \frac{(\omega L)^2}{R} = \frac{(2\pi(4000)(1.5\times10^{-3}))^2}{.126} = 1.12\times10^{4}\,\Omega \tag{32}$$

Usually, most power amplifiers have a lower output impedence. If we connect the resonated coil to the primary of a coil having a turn ratio of 1:15, a power amplifier will see a load of 49.8Ω and will require 22.3 vrms to produce the ten-watt output power.

The above calculations are for the coil in air. In water, the resistance of the coil increases according to equation (21), which for our case amounts to an increase of 50% to a resistance of .189Ω. Also the inductive reactance decreases by an amount, ΔX, given by (reference (f), appendix C)

$$\Delta X = -\frac{2\pi}{3\sigma a}\left(\frac{a}{\delta}\right)^5 \tag{33}$$

which, for our coil, is totally negligible being about 2×10^{-6} ohms. Thus, the main effect of immersing the coil in water is to reduce its Q to about 2/3 of its in-air value.

At 40 Hz, the coil's reactance drops by a factor of 100 and we have Q=3. With such a low value of Q, the concept of resonance is rather ill defined since the value of shunt capacitance, which minimizes the input current, is different from that which makes the impedence real. In any event, the procedure for driving the coil is similar to that at 4 KHz. Additionally at 40 Hz, there is essentially no change in the coil's Q when immersed, since at that frequency the coil is very small compared to a skin depth.

4.3 The Filled Solenoid as a Source

Thus far we have been considering the planar coil as our source. It is also possible to use a permeable material as the source. The idea is to excite the material to a magnetized state, the dipole moment being the volume integral of the magnetization. We shall consider solenoids, long compared to their diameter, filled with a permeable material.

In considering this approach, one immediately encounters two loss mechanisms not present in air-wound coils. One is eddy currents which are set up in the core by the magnetizing field. These can be minimized, in the case of a conductive core, by laminating or powdering the metal, or by using nonconductive types of material such as ferrite. The hysterisis losses are characteristic of the material. Both loss mechanisms are more significant at higher frequency, making the design of a 4000 Hz filled source less efficient than one at 40 Hz.

At 40 Hz for Selectron, the manufacturer's tables (reference (j)) can be interpolated to show a core loss (eddy current plus hysterisis) of .13 Watts per pound at 40 Hz for B = 10 Kilogauss = $1 Wb/m^2$. From the manufacturer's magnetization curve the H required to reach this B is H=.35 oersteds= $27.9 \text{ amp } m^2 \cdot m^{-3}$
We calculate the magnetization, M, which is the dipole moment per unit volume according to

$$M = \frac{1}{\mu_0} B - H \tag{34}$$

which for the above values of B and H yields $M = 7.96 \times 10^5 \frac{\text{amp } m^2}{m^3}$. The volume, V, of Selectron required to equal the 67.9 amp m^2 moment of the 10-watt, 10-pound copper coil is

$$V = \frac{m}{M} = \frac{67.9}{7.96 \times 10^5} = 85.3 \text{ cm}^3 \tag{35}$$

Since the density of Selectron is 7.65 g/cc, we need only 1.44 pounds to produce the same moment as our 10-pound copper coil. Using the figure of .13 W/pound, we see that only .19 Watt is dissipated by core heating.

Now we need to calculate the weight and power needed in the core windings to produce the above magnetization. To do this, we need to specify the rod's shape. For convenience, let the rod's length, l, be 10 times its diameter, d. For a volume of 85.3 cm^3, this means a rod 8.72 inches long with a .872 inch diameter.

For any material, an applied field H_o is required to produce a field H. These are not the same because of the internal demagnetizing field. They are, however, related according to (reference (k))

$$H_o = \frac{N_D}{4\pi} B + \left(1 - \frac{N_D}{4\pi}\right) H \tag{36}$$

The above equation is in Gaussian units and N_D is known as the demagnetizing factor, which depends on the core shape. For a circular rod of 10:1 aspect ratio, $N_D/4\pi = 1.72 \times 10^{-2}$ (reference (k)). For our values of H and B, we obtain H_o=172 oersteds.

I have previously (reference (l)) calculated the magnetic field H_o produced by a solenoidal winding of power-mass product PM as

$$H_c = \frac{1}{\pi \lambda d} \sqrt{\frac{PM}{\rho \rho_m}} \tag{37}$$

in rationalized MKS units. Solving for PM we find PM=6.83 Watt Kg=15 Watt-pound pounds. We can split this up between power and mass in any way we desire. Choosing P=9.5 Watts we find M=1.6 pounds. This corresponds to a winding along the core which is roughly 6mm thick.

Comparing the filled solenoid with the coil at 40 Hz, the filled solenoid (with windings) is 8.72 inches long with a 1.34 inch diameter consuming 9.7 Watts and weighting 3.04 pounds. The coil weighs 10 pounds, consumes 10 Watts, is .5 meters (19.7 inches) in diameter, and has windings .9 inch in diameter. Thus the filled solenoid is more compact and weighs less.

At 4000 Hz it initially is not clear whether a filled solenoid can be designed which is more efficient than a coil. Even 1 mil Selectron would be too lossy requiring a 4.2 pound core and dissipating over 40 Watts in core losses to produce the same magnetic dipole moment as the coil. Bill Bangston of Arnold Engineering stated that their powdered iron cores would be too lossy at 5KHz and suggested ferrites. Large ferrite cores are available, on special order, from Indiana General. For transmitter purposes, their 05 premium material, which is similar to the Ferroxcube 3C8 material, is near optimum. We shall use the material constants for 3C8 (since I don't have an Indiana General catalogue yet).

Running through the same calculations as before, we find we can obtain a dipole moment of 55.7 amp $-m^2$ at 4000 Hz with a 3C8 core 12.2 inches long and 1.2 inches in diameter. It will weigh 2.41 pounds and dissipate 5 Watts. The copper windings will weigh 1.07 pounds and dissipate 5 Watts also. Thus, using the same power, we can obtain the same magnetic moment as the 10-pound coil but with a weight of 3.5 pounds and a maximum dimension of one foot. We have not included seawater losses for the solenoid, whereas we did for the coil. These losses may be smaller than for the coil because the solenoid is physically smaller (see equation 21).

5.0 RECEIVING CONSIDERATIONS

5.1 The Matched Receiver

It is reasonable to expect that the optimum receiver input network combination is where the receiver, be it coil or solenoid, is impedence matched to the input amplifier so as to yield maximum power transfer. This can be accomplished by putting a capacitor in series with the sensor to cancel its inductance and then placing a matching transformer across this combination to make its internal resistance equal to the amplifier's input impendence. If this is done, the voltage, V_m, appearing across the input terminals of the amplifier is

$$V_m^2 = \frac{V_o^2 R_i}{4R} \tag{38}$$

where V_o is the open circuit sensor voltage, R_i the input impedence of the amplifier and R the internal resistance of the coil or solenoid (winding resistance plus any series loss resistance due to core losses).

The equivalent noise input, Vn, across the amplifier is given by

$$V_n^2 = 4KT(\Delta f) R_i F \tag{39}$$

where K is Boltzmann's constant, T the absolute temperature, Δf the bandwidth, F a factor which is unity for an ideal amplifier, and >1 for real amplifiers. The signal-to-noise power ratio is given by

$$SNR = \frac{V_o^2}{RF} \frac{1}{(16KT\Delta f)} \tag{40}$$

Maximizing the signal-to-noise ratio for a given amplifier is equivalent to optimizing the quantity V_o^2/R for our sensor. Letting R be due solely to the resistance of the windings, we find for the coil

$$\sqrt{\frac{V_o^2}{R}} = \left[\frac{a}{2}\sqrt{\frac{M}{\rho\rho_m}}\right]\omega B \tag{41}$$

and for the solenoid

$$\sqrt{\frac{V_o^2}{R}} = \left[\frac{a}{2}\frac{4\pi}{N_D}\sqrt{\frac{M}{\rho\rho_m}}\right]\omega B \tag{42}$$

where in each case a is the radius, M the mass of the windings, and N_D the demagnetization factor.

Comparing equation (41) to equation (22), we see that the same factor appears in both. Had we continued on with equation (37) making the assumptions of a linear material, no core loss and no heating of the sea water, we would have found for the filled solenoid

$$m = \left[\frac{a}{2}\frac{4\pi}{N_D}\sqrt{\frac{M}{\rho\rho_m}}\right]\sqrt{P} \tag{43}$$

which is very similar to (42). Thus both for the coil and the solenoid, the same parameters optimize the sensor, whether as a transmitter or receiver , if we neglect core losses and water heating. This is reasonable from reciprocity. It is interesting that the receiving equations contain a factor of ω not present in the transmitting equations. This is because we optimized the transmitting equations for the near field. Had we optimized the transmitter to give the largest value of H in the far field in air, then the transmitting equations would also contain a factor of ω and we would have complete symmetry.

It is interesting to calculate the receiving response at 40 Hz for the filled solenoid of the last section. Assuming an ideal amplifier of 12Ω input impedence, we find from equation (39)

$$V_n = \sqrt{(4)(1.38\times10^{-23})(295)(1)(12)} = .442 \ nV \tag{44}$$

for a 1 Hz bandwidth at room temperature. From equation (42) for our solenoid of section 4.3 we find $(V_i^2/R) = 22.08\,\omega B$. For $B = 1\gamma = 10^{-9}\ wb/m^2$
We find from equation (38)

$$V_m = 9.61\ \mu V \tag{45}$$

The signal-to-noise ratio is then

$$SNR = 20 \log \frac{V_m}{V_n} = 86.7\ dB \tag{46}$$

5.2 Direct Receiver Connection

One of the problems with the matched receiver is that the phase of the signal is sensitive to any slight change in the impedance of the elements. Another problem is the weight of the matching transformer. Both these problems can be eliminated by connecting the sensor directly to the amplifier. If this is done, then the voltage across the amplifier is

$$V_m^2 = \frac{V_i^2}{(R/R_i + 1)^2 + (\omega L/R_i)^2} \tag{47}$$

Usually we will have $R_i >> R$. In this case, the maximum signal will be obtained with a turn ratio such that $\omega L = R_i$. However, with this turns ratio, the reactive and resistive components are equal and the currents phase is still sensitive to the fluctuations in the sensor's impedance. To eliminate this sensitivity, we must make $\omega L << R_i$.

A solenoid of length, l, radius, a, and permeability, μ, will have an inductance, L, given by

$$L = \frac{\pi a^2 \mu_c N^2}{\ell \left(\frac{N_o}{4\pi} + \frac{\mu_c}{\mu} \right)} \tag{48}$$

For Selectron $\frac{N_o}{4\pi} >> \frac{\mu_c}{\mu}$ so that the inductance is determined by the core's shape rather than its permeability. If we assume an input resistance $R_i = 12\Omega$ and require $\omega L/R_i = .1$, then we find N=187 for the solenoid of section 4.3. For single layer coverage of a 8.16 inch long core this would roughly correspond to 43 feet of #17 wire which would yield a winding resistance of .218Ω and a weight of .27 pounds. The winding's weight is considerably less than the 1.6 pounds of the last section.

For R and ωL both less than R_i, we have

$$V_m \cong V_o \tag{49}$$

For a solenoid

$$V_o = \pi a^2 N \omega \frac{4\pi}{N_o} B_o \tag{50}$$

Plugging in the numbers we find that for $B_o = 1\gamma$, f=40 Hz, $V_o = 1.07\mu V/\gamma$. This represents a 19.1 dB drop in performance from the ideal case. Considering that we will be using a real amplifier with F>1 the signal-to-noise ratio will drop even more. However, considering the large ideal SNR, this lower performance is no real handicap and is more than made up for by the simple receiving arrangement which requires no capacitor or transformer and is insensitive to impedence variations.

Another benefit in the above arrangement is that the core can be quite lossy at 40 Hz. Even if $Q=\omega L/R$ is on the order of unity, the output voltage will still be V_o. At 4000 Hz, the situation changes, since it is impossible to make $\omega L/R_i=.1$ for $R_i=12$ ohms without going to a very heavy winding or a loosely spaced winding, both of which are probably undesirable. Making $\omega L/R_i$ small by increasing R_i lowers the signal-to-noise ratio (for F constant) since the noise increases with R_i. If the sensor is to be directly connected at 4000 Hz, it is probably optimal to make $\omega L>>R_i$. This also requires $\omega L>>R_i$ to make the current's phase insensitive to impedence variations.

It appears that the Indiana General Q1 antenna core, 7.52 inches long by .620 inch diameter will make a suitable receiving antenna at 4000 Hz (and with a different winding, also at 40 Hz). Letting $\omega L=10R_i$ at 4000 Hz implies about 250 turns of #22 wire; a winding resistance of $.655\Omega$. We estimate a core loss resistance of $.6\Omega$ and a very rough estimate of seawater resistance of 1.4Ω for a total of $R=2.65\Omega\Omega$and $Q\approx45$ in water. V_o for this coil will be $65\mu V$, and Vm will be $.65\mu V$.

6.0 CHOICE OF OPERATING FREQUENCY

The previous examples were based on using 40 Hz and 4000 Hz as the operating frequencies. 4000 Hz was chosen as an upper limit for consideration since here attenuation length is 4 meters, which is a tolerable working height. Beyond this frequency at 4 meters, the magnitude of the effect decreases significantly due to attenuation. 40 Hz seems to represent a convenient experimental lower limit for lab work. However, the choice of optimum frequency cannot be determined from theoretical considerations alone and must be made primarily on the basis of experimental work. We can, however, enumerate the tradeoffs involved.

Since there are two effects, the magnetic and eddy current involved, probably the most important factor is to determine that frequency which has the largest net effect. This can be determined fairly easily in the lab by placing a coil in proximity to a material sample representative of the pipeline and noting the fractional difference between the coil's inductance in free space and when isolated. This fractional change is proportional to the real effect produced by the pipeline. Noting the fractional change as a function of frequency will determine the frequency where the effect is maximum.

Although the magnitude of the effect at the operating frequency is a major consideration, other factors may be equally important. At 40 Hz, using a solenoid results in a very compact source. At this frequency the magnitude of the eddy currents induced in the vehicle, which are a potential source of interference to pipeline detection, should be less than at higher frequencies. 40 Hz also has the advantage of having a 40-meter skin depth in water so that its performance in air and in water should be identical, facilitating the transfer of laboratory results to the operating environment.

Higher frequencies also have advantages. At 4000 Hz the signal returning from the pipeline 4 meters away will be mainly in quadrature with the signal arriving directly from the transmitter coil or other nearby objects. Signal-detection may be enhanced by concentration on quadrature detection. The effect is not available in air nor in water at 40 Hz because the desired signal undergoes no phase change relative to the interfering signals in these situations. True noise, arising from the free swimmer motors or filtering in from the air-sea interface, should also be minimized, the former because less noise is probably generated at the higher frequency, and the latter since an attenuation length at 4 KHz is only 4 meters. Finally the theoretical matched receiver output voltage (see section 5.1) increases linearly with frequency. This advantage, however does not carry over to the directly connected case.

7.0 ELIMINATION OF THE BACKGROUND FIELD

7.1 Introduction

Thus far we have looked at the problem as if we were trying to detect the magnetic field from the pipeline in the absence of other factors except possibly receiver noise. In practice the signal impinging on the receiver which comes directly from the transmitter will be orders of magnitude larger than the induced pipeline signal. While it is true that a constant background signal, however strong, does not interfere with detection, the interfering signals, both those coming from directly from the transmitter and indirectly via induced currents on the vehicle, will not remain constant. Variations in transmitter power, thermal effects on electronics and mechanical structures and vibrations, among other effects, will cause the background to vary. Even small variations of this large background will tend to mask the desired pipeline signal, being especially serious if the temporal variations are similar in the desired and background signal.

To give an example of the magnitude of this effect, consider the 40 Hz Selectron source at 4 meters from the pipeline. The returned signal is about $4\gamma = 4 \times 10^{-5}$ gauss. At the immediate vicinity of the core, the magnetic field coming directly from the core is equal tothe core's induction of 1×10^{4} gauss, a signal to background ratio of 4×10^{-9}. Clearly, even very small fluctuations in the background signal are intolerable.

In the following, we will examine schemes which discriminate against the direct field. Employment of these schemes will usually also discriminate against the desired signal. Such a loss is tolerable provided we are well above the system and ambient noise level. A useful scheme is one which discriminates against the direct signal to a greater extent than the pipeline signal is discriminated against.

Schemes for direct field discrimination fall into two categories, predetection and post detection. Predetection schemes are those which operate before the sensor detects the field, post detection schemes operate after the magnetic field is translated into an electrical signal. In general, predetection schemes are preferred in that they are not influenced by electrical instatilities. However, post detection schemes will probably be necessary since it is doubtful that enough discrimination will be available from predetection schemes alone.

7.2 Predetection Background Elimination

7.2.1 Source Receiver Séparation. By moving the source or receiver toward the pipeline, the background decreases while at the same time increasing the desired effect due to the pipeline. Of course this directly decreases standoff distance. For a constant standoff distance, the background can still be discriminated against by moving the receiver away from the transmitter anywhere in the plane perpendicular to the one joining the free swimmer and pipeline. This separation discriminates against the direct signal without appreciably affecting the pipeline signal until the separation becomes on the order of half the distance to the pipeline.

To see how the background is reduced by separation, consider the receiver separated from the 40 Hz Selectron solenoid by one meter. Along the transmitter's axis the magnetic field is

$$B = \frac{\mu_0 m}{2\pi r^3} \tag{51}$$

For m=67.9 amp m^2 and r=1 meter we find B=1.35 x $10^4 \gamma$. This is a factor of 10^5 less than the field existing in the transmitter's core.

7.2.2 Shielding. The geometry of the paths taken by the pipeline signal and the direct background are quite different. By incorporating some shielding material in the region between the source and receiver, it may be possible to discriminate against the direct signal in favor of the pipeline signal. Either conductive materials, producing eddy currents or magnetic materials, becoming magnetized, might produce the required discrimination. The exact signal to background improvement to be obtained from shielding is difficult to calculate in advance, and is best determined from laboratory experiments.

7.2.3 Orthogonality of Source and Receiver. Both the magnetic field and our sensors are vector in character. This means we can align our receiver so that their axes are perpendicular to the field coming from the transmitter. The direct magnetic field can then, in principle, be nulled completely. There will also be a reduction in the pipeline signal. If the source and receiver are separated on the order of the distance to the pipeline, the pipeline signal will not be attenuated as much as the direct signal.

In practice obtaining a perfect null will be impossible, probably mainly due to mechanical vibrations which keep the receiver from being aligned exactly perpendicular to the incoming direct field. For example, for the 8.72 Selectron rod clamped at its center, the ends must be positioned to within 1 milimeter to reduce the detected signal to 1/100 of its maximum value. For a receiver located 1 meter from the 40 Hz Selectron source, this would reduce the effective direct field to 135γ.

7.3 Post Detection Processing

7.3.1 Differencing. If the free swimmer is to follow a pipeline, it is important to obtain control information whenever the free swimmer wanders from its desired position directly over the pipeline. A convenient way of accomplishing this is to use two receivers laterally displaced from the transmitter but symmetrically placed with respect to it. These receivers are then differentially connected so that when the free swimmer is directly over the pipeline the pipeline produces no net signal. When the free swimmer wanders from directly above the pipeline, a signal is produced with a polarity corresponding to that of the receiver which is closest to the pipeline. The strength and polarity of this signal can then be used as the control signal to keep the free swimmer over the pipeline.

If the two receivers are placed symmetrically with respect to the transmitter, we would expect that each receiver will pick up the approximately same amount of direct field. When they are connected differentially, this direct background will cancel. This cancellation can be made more perfect by putting a potentiometer in series with one receiver and adjusting it for deepest null. Only lab experiments will be able to determine the depth of this null. If 40 dB can be obtained, then when used in conjunction with 1-meter source receiver separation and orthogonal receiver orientation, the direct signal pickup will be equivalent to 1.4γ for the 40 Hz Selectron source. This would put the direct background near or below the level of the signal from the pipeline four meters away, and we should begin to be able to pick out the pipeline signal.

As for the predetection schemes, differencing also discriminates against the desired pipeline signal. The degree of discrimination is dependent on where the pipeline is located relative to the two receivers. For example, when the pipeline is centered between the two receivers, the pipeline signal is opposite and nearly equal so that there is complete cancellation of the pipeline signal. At other positions of the pipeline, the cancellation is less but still finite. The largest net signal will occur when the pipeline is centered halfway between the transmitter and one of the receivers. For the case of the transmitter separated from the receiver by half the depth of the pipeline, this maximum is about half the pipeline signal in any one of the receivers.

7.3.2 Adaptive Differencing. In the above scheme, the null was obtained by setting a potentiometer. In the course of a mission this null point can be expected to drift due to thermal and mechanical changes. The system will have to be re-nulled, perhaps periodically. As long as the system is essentially drift free for fairly long periods of time, this will cause no problem since auto null circuits can be energized periodically when the vehicle is away from the pipeline. At 40 Hz, it appears only an amplitude adjustment will be necessary. At 4000 Hz both amplitude and phase compensation may be necessary. This last case is very similar to adaptive noise cancelling of a single sinasoid and is known to require a two-tap delay line with two multipliers.

7.3.3 Phase Effects. In air the wavelength at 40 or 4000 Hz is so much greater than any of the distances involved that the phase of the received signal is the same as the transmitted signal. The same situation obtains in water at 40 Hz where the wavelength, which is 2 times the skin depth, is over 250 meters. However, at 4000 Hz the wavelength is only 25 meters and the pipeline signal's phase at the receiver can be considerably different from either the transmitted phase or the direct signal from the transmitter. For example, at 4 meters height, the pipeline signal is in quadrature with the transmitter and nearly in quadrature with the direct signal. By operating the reference signal of our lock-in amplifier at a phase which is in quadrature with the direct signal, we can again discreminate against the direct signal in favor of the pipeline signal. Obtaining a reference which is phase locked to the direct signal is easy in the absence of a target, since this is just the received signal.

The exact implementation and gains to be obtained by using phase discrimination are not clear, although perhaps very significant for a 4000 Hz system. Unfortunately, experiments involving phase effects can only be done under water. Studies of phase effects should probably be postponed until in-air experiments are complete. If it eventually appears that a 40 Hz system will suffice, there will be no reason to investigate phase effects at all.

8.0 SUGGESTIONS FOR EXPERIMENTS

Table 1 shows the type of equipment necessary to develop a magnetic pipeline detection system. These have all been mentioned in the text previously, and are all currently on order or in-house. Another requisite for lab experiments, especially for simulation of actual pipeline detection, is a large clear area free from other metals which may become magnetized or produce eddy currents.

A suggested plan for performing the magnetic experiments will now be given. First the transmitting and receiving coils and solenoids should be wound as described in sections 4.1 through 4.3 and section 5.2. Their complex impedence should be measured as a function of frequency and compared with the formulae given in the test. Since winding capacitance and eddy currents were not taken into consideration in the test formulae, the transmitters and receivers may have to be rewound. In general the measured Qs will be less that those calculated, since the effects we've ignored tend to lower inductance and increase resistance. It must be determined whether this lower Q is sufficiently significant to make the subject solenoid or coil too inefficient to be used as a transmitter and receiver.

Once the basic characteristics of the transmitters and receivers have been determined, their performance in conjunction with the appropriate driving or receiving electronics should be checked. The minimum detectable signal and dynamic range should be checked for the sensor-preamp-locking combination. The transmitters should be driven to their full power output where efficiency and magnetic moment measurements should be made.

The range of suitable operating frequencies should be examined by measuring the change of inductance vs. frequency with various magnetic and nonmagnetic materials in proximity of a transmitting coil. Frequencies near where the eddy current and magnetic effects cancel will produce little effect and should be avoided.

Using a single source and receiver, experiments should be conducted to see what depth of null can be maintined for an orthogonal receiver in the presence of thermal and mechanical fluctuations. Next, a separate receiver, identical and placed symmetric with the first, should be used to test the depth and quality of the null obtained with a differentially connected pair. Conducting and magnetic materials should be placed around the receivers to see if shielding can improve the discrimination of the direct signal.

If all looks well at this point, laboratory pipeline detection experiments should begin. Two receivers mounted on a nonmagnetic beam, separated by 1 to 2 meters, with a transmitter in the center can be used to detect a pipe. Measurement of the detected signal versus range and lateral offset can then be made. Comparison can be made with equation (13) to obtain an estimate of the system's performance against different sized pipes in sea water. After the system has been tested in isolation, it should be mounted on the vehicle in the laboratory and tested to see that proximity to the vehicle has not degraded the system.

The last experimental phase would be in-water tests. Two receivers and a transmitter could be waterproofed, mounted on floats and towed behind a small boat over suitable sections of pipe in San Diego bay. Pipe detectability for different sized pipes at different depths could be determined as a function of pipe diameter and depth. Also salt water experiments are the only ones where quadrature effects could be investigated. Demonstration of the ability to detect pipelines with a towed system would probably justify continuance of the program to incorporate pipeline inspection in free swimmer.

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